

MASSACHUSETTS INSTITUTE OF TECHNOLOGY
LINCOLN LABORATORY

POLARIMETRIC SAR ANTENNA CHARACTERIZATION

R.M. BARNES
D.J. BLEJER
Group 47

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ABSTRACT

A procedure is described for obtaining the two-way polarimetric properties of a SAR antenna from one-way measurements in a compact antenna test range. The two-way properties are determined by invoking reciprocity and computing a polarization distortion matrix (PDM). Next, one-way antenna-only measurements are combined with boresight measurements of the complete system to find an overall PDM. Polarimetric distortions caused by the antenna can be compensated using the inverse of the PDM; however, the Advanced Detection Technology Sensor antenna was found to have adequately low crosspolarization without the need for crosspolarization compensation.

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1. INTRODUCTION

Lincoln Laboratory has built a high resolution, polarimetric, synthetic aperture radar under DARPA sponsorship. This radar, called the Advanced Detection Technology Sensor (ADTS), has been used to collect high quality data on clutter and on stationary targets in clutter. The data will be used to evaluate the performance of polarimetric detection and classification algorithms under the ADT program.

An important system component that significantly affects radar polarimetric quality is the antenna. The radar uses an Alpha Industries model A858-12 scalar-horn/lens antenna that scans in azimuth and elevation inside a monolithic spherical radome. Figure 1-1 shows a drawing of the antenna, gimbal, and radome. Since a scalar horn antenna supports the HE_{11} mode, which is highly linearly polarized, the antenna was known to have a high degree of polarimetric purity. The polarimetric distortion that the radome might introduce was not known.

A measurement program was established to characterize the antenna/radome properties over the radar bandwidth ($33.6 \text{ GHz} \pm 300 \text{ MHz}$). The ADTS two-way system specification requires that

The cross-polarization isolation of the sensor and the processor shall be at least 30 dB along the antenna boresight and at least 20 dB within the copolarized 3 dB beamwidth; this result shall hold after calibration/compensation.

This is a two-way specification that applies after the signal has been transmitted and then received from a target.

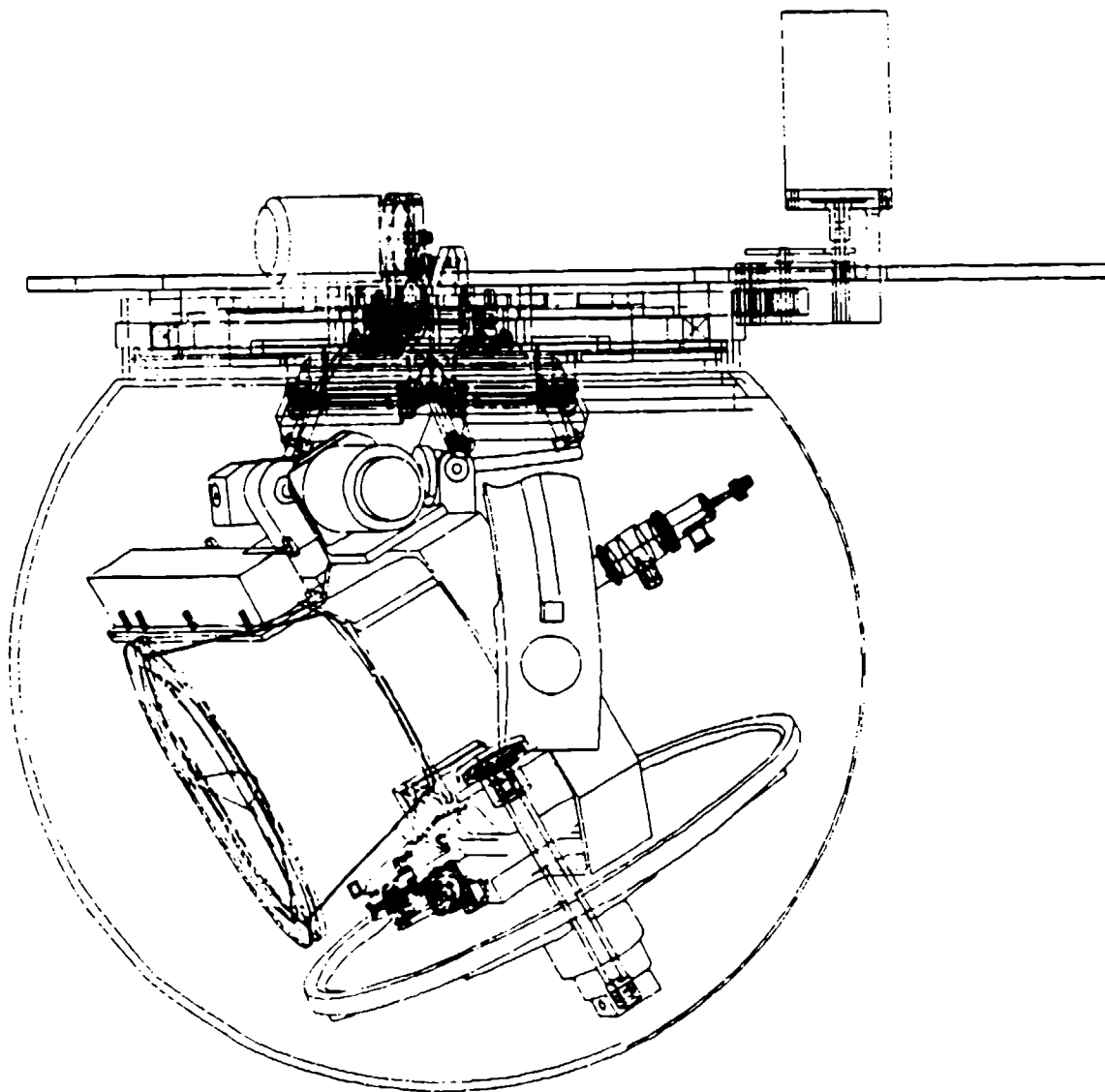
The purpose of this report is to document the techniques used to calibrate the ADTS antenna when used as part of the radar system. In particular, the methods used to combine antenna-only measurements in a Compact Range and total-system measurements with the aircraft are described.

Section 2 describes our measurements of the one-way properties of the ADTS antenna, including (1) cuts in both cardinal and intercardinal planes, and (2) contour plots, of both the copolarized and crosspolarized one-way patterns.

Section 3 explains how the measured one-way properties of the antenna can be used to predict the two-way properties.

Section 3.1 begins with the definition of one-way and two-way polarization distortion matrices. The two-way properties of the antenna are determined by invoking reciprocity and computing a polarization distortion matrix (PDM).

Section 3.2 shows how we can use the distortion data to take corrective action, if the two-way distortion is not within specification: polarimetric distortions caused by the antenna are compensated by (1) forming a complex vector from the elements of a measured polarization



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Figure 1-1. ADTS antenna gimbal and radome..

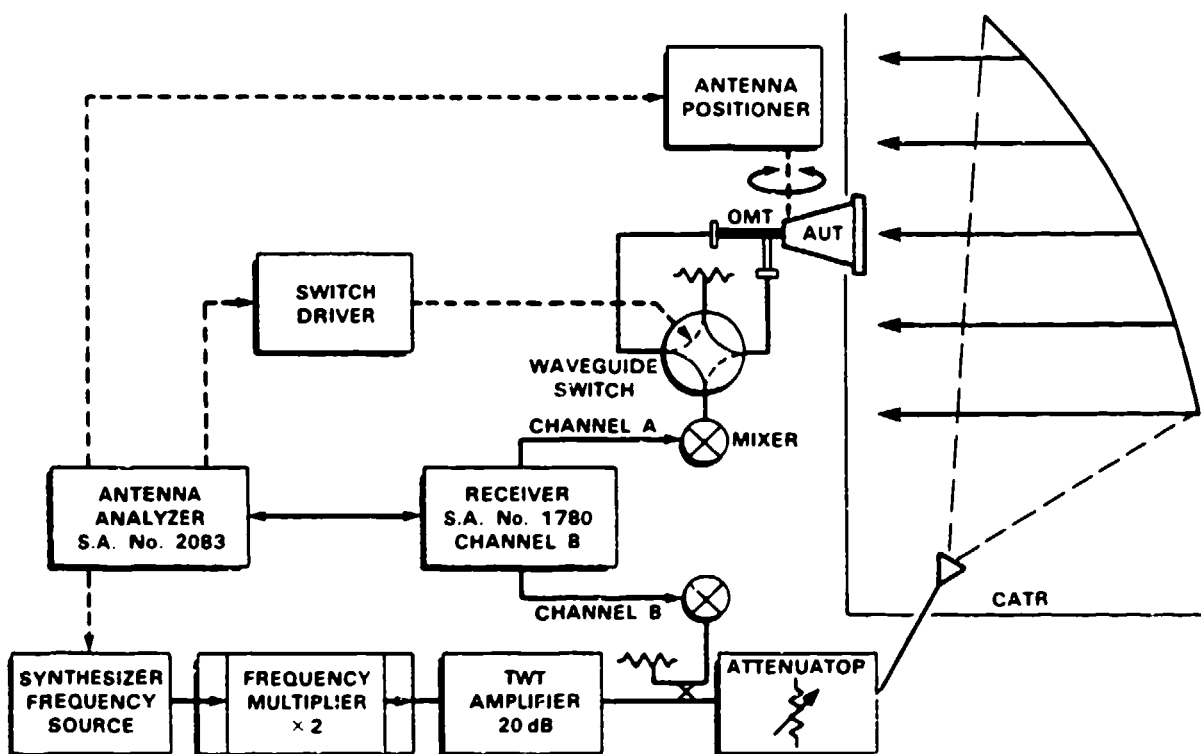
scattering matrix (PSM), (2) premultiplying by the inverse of the PDM, and (3) forming an estimate of the true PSM of the target from the resulting complex vector.

Section 3.3 describes how we augment the antenna-only measurements of section 2 with complete system measurements, including transmit and receive waveguide components.

Section 4 shows that the worst-case two-way crosspolarization is 6 dB worse than the worst-case one-way crosspolarization; this result is interpreted for the example of the ADTS antenna.

2. ONE-WAY PROPERTIES OF THE ADTS ANTENNA

The one-way properties of the ADTS antenna with and without the radome were measured at a compact antenna test range at Lincoln Laboratory. This range uses a parabolic reflector and instrumentation from Scientific-Atlanta. A block diagram of the measurement system is shown in Figure 2-1. A small feed horn illuminates the reflector, which produces a plane-wave field in the vicinity of the antenna under test (AUT). The AUT is rotated in azimuth and elevation to obtain a receive pattern. By the Reciprocity Theorem, the transmit pattern of the AUT is identical to this receive pattern.



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Figure 2-1. Compact Antenna Test Range (CATR) system block diagram..

One-way measurements of amplitude and phase for all polarizations were taken at the radar center frequency and at the upper and lower frequencies which define the radar bandwidth. Principal plane patterns and raster scan measurements were taken. To emulate the actual

ADTS radome, a 14 inch diameter spherical section of acrylic was fabricated to have the same thickness, material properties, and curvature as the actual radome.

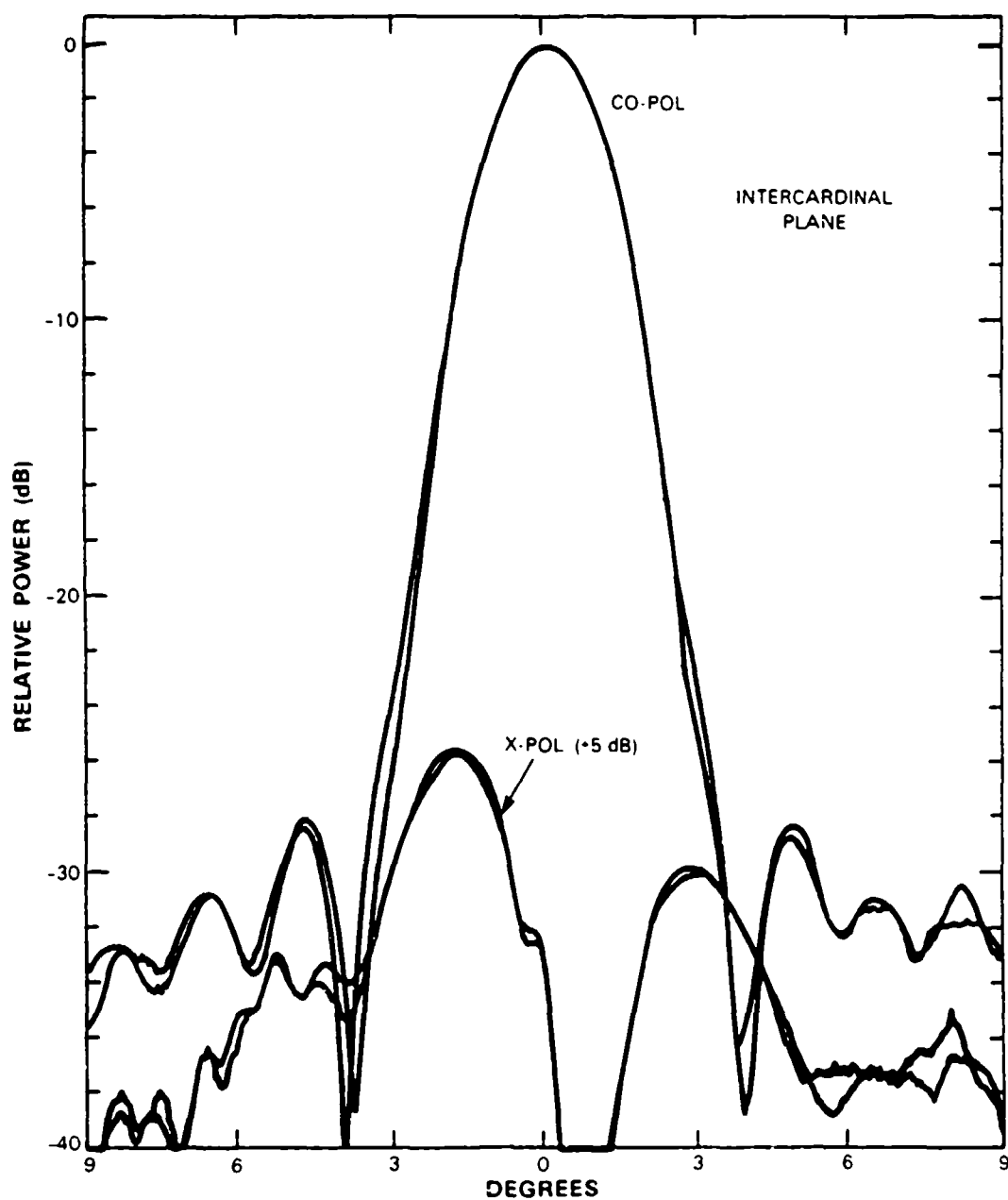
Figure 2-2 shows co-polarized and cross-polarized amplitude patterns in an intercardinal plane at the center frequency, 33.6 GHz. The patterns with and without the acrylic section used for simulating the radome are shown. The co-polarized patterns are essentially identical over the main lobe. Cross-polarization appears mainly in the form of sidelobes in the intercardinal plane with a relative minimum near boresight. The radome does not induce any significant cross-polarization; this is in agreement with theoretical results.

The results for the H-plane, shown in Figure 2-3, are similar to the results for the intercardinal plane. An important difference is that cross-polarization appears as a main lobe on boresight. Theoretically, there is no cross-polarization on boresight for a linearly polarized aperture. Therefore, the observed cross-polarization must be due to imperfections in the antenna, radome, and/or the compact range instrumentation.

Figure 2-4 shows the amplitude pattern for the antenna with the acrylic section. The pattern appears as a contour plot versus azimuth and elevation. The contour plot shows that the main lobe has a high degree of circular symmetry; this is necessary for good performance with circular polarization.

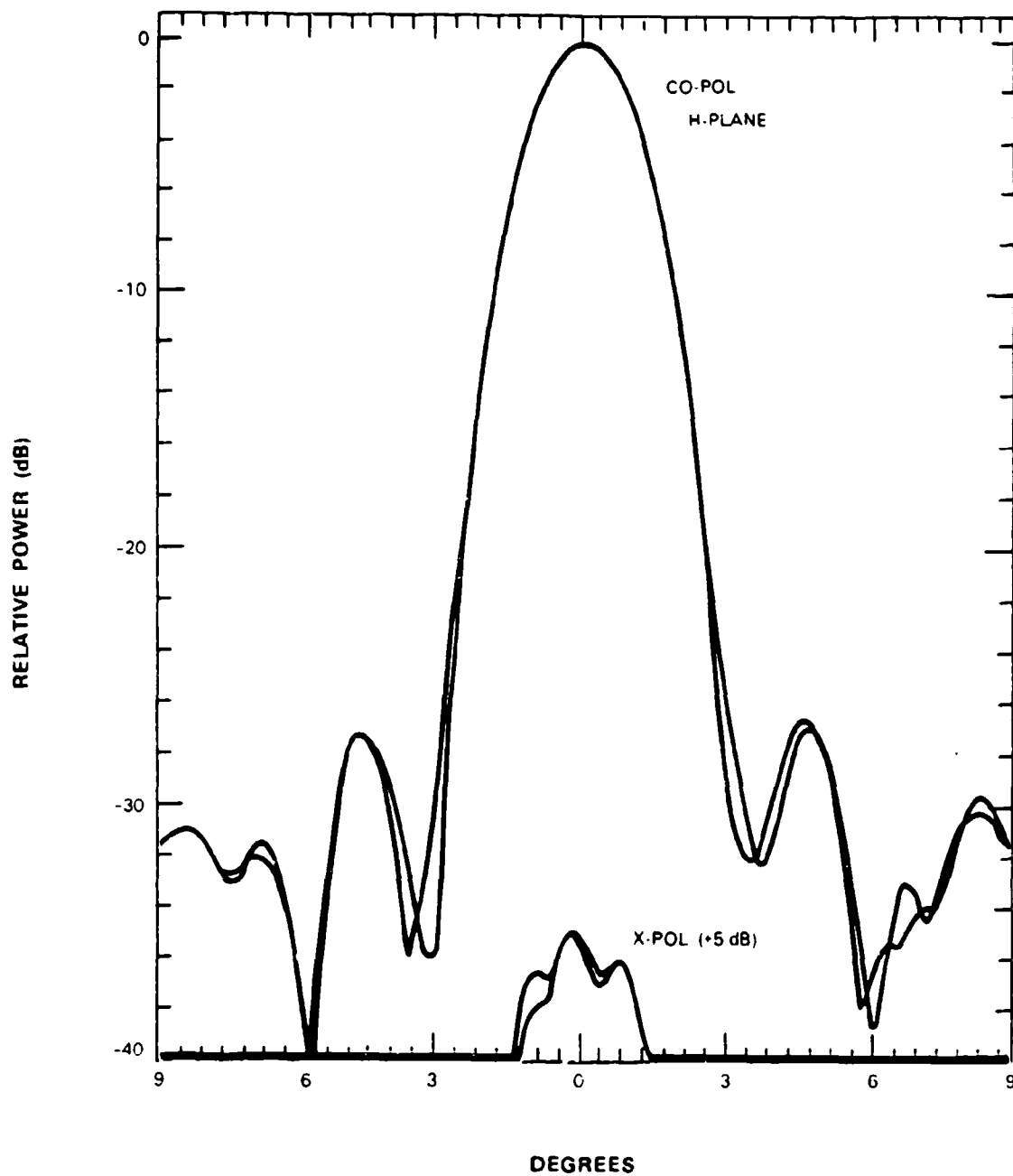
An important parameter that characterizes the polarimetric purity of an antenna is the maximum cross-polarization ratio. The cross-polarization ratio is defined as the ratio of the cross-polarized component to the co-polarized component, at the same azimuth and elevation. Figure 2-5 shows the maximum cross-polarization ratio, over the radar bandwidth, for the antenna with the acrylic section. The maximum (worst-case) cross-polarization ratio is -37 dB on boresight and is approximately -29.5 dB everywhere within the -3 dB main beam.

These results indicate that the polarimetric properties of the antenna with the radome meet the ADTS cross-polarization specification one-way. Section 4 will use theoretical considerations to demonstrate that the antenna with radome meets the specification two-way.



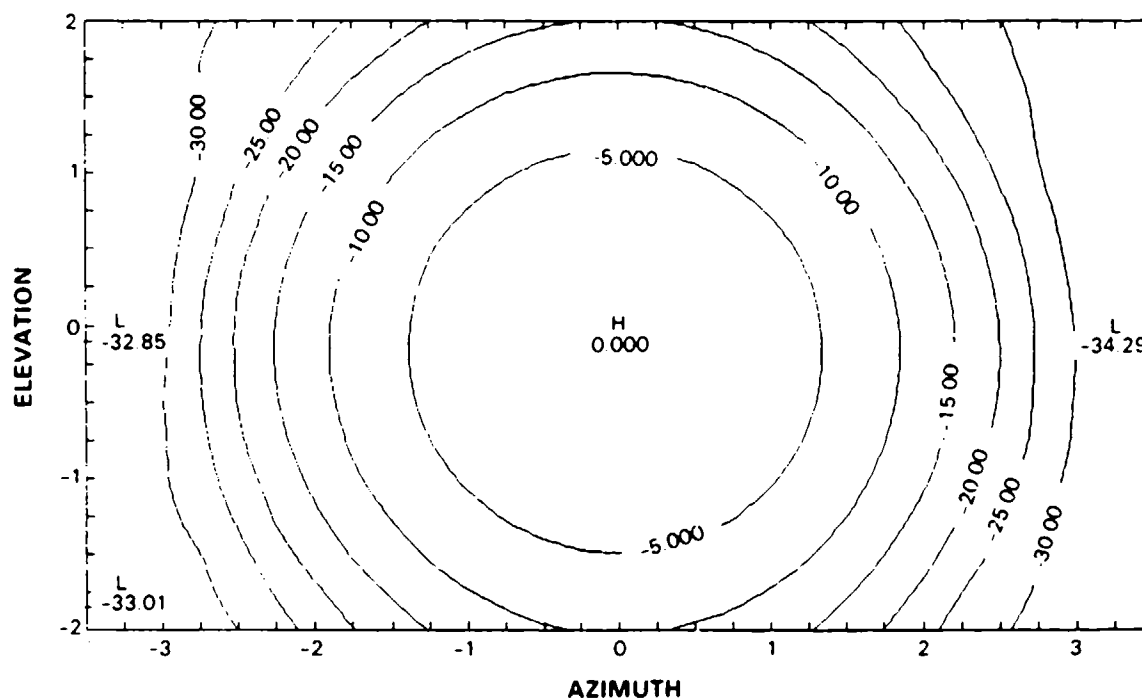
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Figure 2-2. ADTS antenna intercardinal plane measurements \pm radome (33.6 GHz)
(The X-pol values are 5 dB below the values indicated on the scale.).



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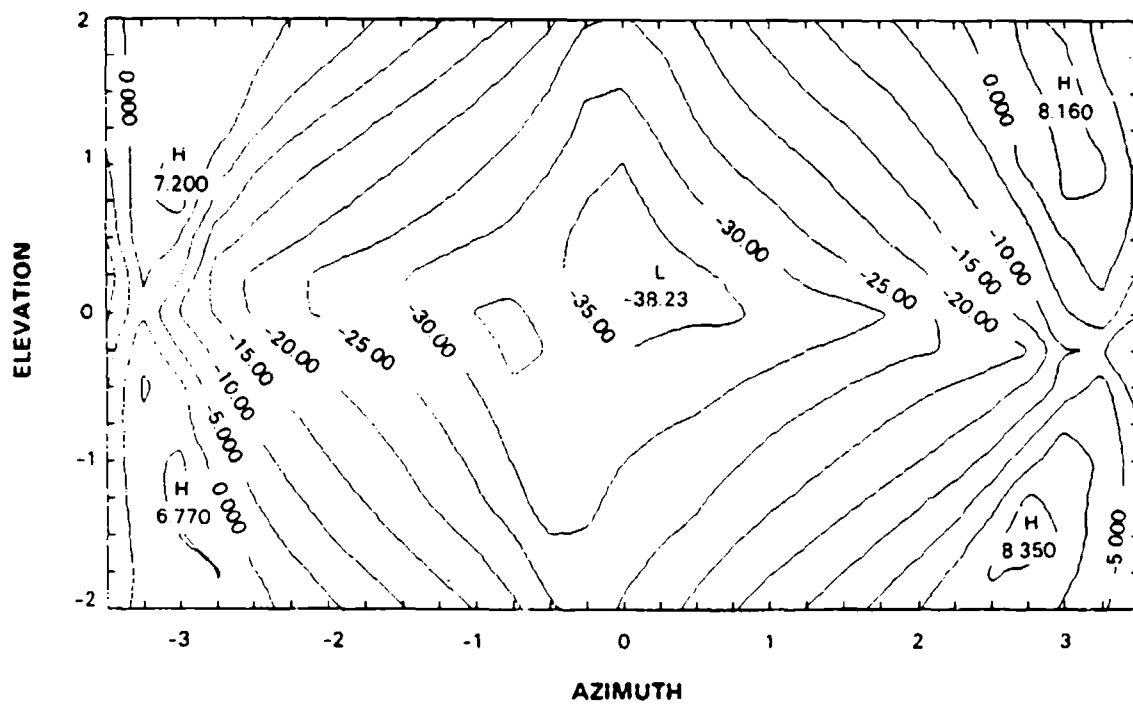
Figure 2-3. ADTS antenna cardinal plane measurements \pm radome (33.6 GHz)
(The X-pol values are 5 dB below the values indicated on the scale.).



AMP (dB) REL TO VV BORESIGHT
 XMT V, RCV V, FREQ = 33.60 GHz
 ADTS1ADIR.DAT
 LINEAR XMIT, LINEAR RECEIVE, W/SHARD, W/SHROUD, W/O POLARIZER
 COPOL

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Figure 2-4. VV power versus azimuth and elevation for 33.6 GHz..



MAX XPOL (dB) REL TO COPOL AT SAME Az, El
 MAXIMUM XPOL = -29.62 dB
 ADTS1ADIR.DAT
 LINEAR XMIT, LINEAR RECEIVE, W/SHARD, W/O POLARIZER

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Figure 2-5. Worst-case HV and VH power versus azimuth and elevation for 33.3, 33.6, and 33.9 GHz.

3. POLARIMETRIC DISTORTION AND CORRECTION

The first step in describing distortion is to define the system for describing the distortion mathematically. Since the notation used in this paper is not widely known, the definitions from an earlier work [1] are repeated here.

3.1 ONE-WAY AND TWO-WAY POLARIZATION DISTORTION MATRICES

The basic idea in polarimetric distortion is the difference between the actual and the measured polarization scattering matrices. The actual PSM is what we would measure if no distortion is present, while the measured PSM can be radically different. We assume here that distortion consists of linear transformations.

Assume that an imperfect antenna actually transmits the vector $\underline{C}_1 = \begin{bmatrix} C_{11} \\ C_{21} \end{bmatrix}$ when it is nominally transmitting horizontal polarization ($\begin{bmatrix} 1 \\ 0 \end{bmatrix}$ in the HV basis). Similarly, assume that the antenna transmits $\underline{C}_2 = \begin{bmatrix} C_{21} \\ C_{22} \end{bmatrix}$ instead of nominal vertical ($\begin{bmatrix} 0 \\ 1 \end{bmatrix}$). Define the complex two-by-two transmit distortion matrix as:

$$C \equiv \begin{bmatrix} \underline{C}_1 & \underline{C}_2 \end{bmatrix} \quad (3.1)$$

When the nominal transmitted field is \underline{t} , the actual distorted transmit field is (by linear superposition) $C\underline{t}$. This matrix describes both channel-to-channel amplitude and phase imbalance (the diagonal terms of C) and cross-polarization leakage (the off-diagonal terms of C).

Denote the actual polarization scattering matrix (PSM) of a target of interest as A . The field scattered from the target back to the receive antenna is $\underline{g} = zAC\underline{t}$. The constant of proportionality z is a complex scalar which embodies amplitude and phase changes due to two-way propagation.

Similar considerations apply to the receive antenna. If the antenna did not distort and the field which it would transmit (if it were an transmitter) is \underline{r} , then the complex voltage measured when the incoming field is \underline{g} would be $V = \underline{r}^T \underline{g}$, where the superscript denotes transposition.

Now consider distortion in the receive antenna. Define the receive distortion matrix B as the distortion matrix which describes how this receive antenna would behave as a transmitting antenna. Then the voltage measured when the incoming field is \underline{g} would be $V = (B\underline{r})^T \underline{g} = \underline{r}^T B^T \underline{g}$.

In general, the transmit antenna and the receive antenna can be physically distinct antennas, and therefore B need not be equal to C .

Combining the definitions of \underline{z} and V , the measured (i.e. distorted) voltage when both transmit and receive antennas may distort is:

$$\begin{aligned} V &= \underline{z}(B_R)^T(AC\underline{t}) \\ &= \underline{z}_R^T B^T AC\underline{t} \end{aligned} \quad (3.2)$$

If the measured (distorted) PSM is denoted as E , then

$$E = \underline{z} B^T A C \quad (3.3)$$

The bookkeeping needed to describe both transmit and receive distortion can be combined in a two-way distortion matrix. One useful way to do this is to rearrange [2] [1] the elements of the PSM A into a complex four-vector \underline{a} , which is called the polarization scattering vector

$$\underline{a} = \text{vec}(A) = \begin{bmatrix} A_{HH} \\ A_{HV} \\ A_{VH} \\ A_{VV} \end{bmatrix} \quad (3.4)$$

A similar vector \underline{e} can be formed from the elements of the matrix E .

It can be shown [3] [1] that the distorted polarization scattering vector \underline{e} is equal to

$$\begin{aligned} \underline{e} &= \underline{z} D \underline{a} \\ \text{where } D &= B^T \otimes C^T \end{aligned} \quad (3.5)$$

The two-way distortion matrix D is the Kronecker product (also known as the direct product) between the transposes of B and C .

Suppose that both the antenna system and also the receiver/transmitter system contribute to distortion. This is expressed mathematically as follows:

$$E = \underline{z} B_R^T B_A^T A C_A C_R \quad (3.6)$$

where C_R and B_R are the transmit and receive distortion matrices, respectively, of the system RF plumbing, and C_A and B_A are the transmit and receive distortion matrices associated with the antenna.

The two-way distortion matrix for this concatenated distortion can be shown to be equal to the product of the individual distortion matrices (D_R, D_A) for the subsystems:

$$\begin{aligned} \underline{\epsilon} &= \underline{z} D \underline{a} \\ \text{where } D &= D_R D_A \\ D_R &= B_R^T \otimes C_R^T \\ D_A &= B_A^T \otimes C_A^T \end{aligned} \quad (3.7)$$

3.2 POLARIZATION COMPENSATION

The previous section described how to measure the distortion caused by an antenna. Suppose we wish to correct a measured PSM E for the effects of polarimetric distortion.

If the distortion matrices are estimated [1] as \hat{C} and \hat{B} , then the compensated PSM \hat{A} is found from the measured PSM E as follows:

$$\hat{A} = (\hat{B}^{-1})^T E \hat{C}^{-1} \quad (3.8)$$

A similar relationship is found for the two-way distortion matrix D . Given estimated one-way transmit and receive distortion matrices \hat{C} and \hat{B} , respectively, we find the estimated two-way distortion matrix \hat{D} as

$$\hat{D} = \hat{B}^T \otimes \hat{C}^T \quad (3.9)$$

so that the compensated polarization scattering vector $\hat{\underline{a}}$ is found from the measured polarization scattering vector as follows:

$$\hat{\underline{a}} = \hat{D}^{-1} \underline{\epsilon} \quad (3.10)$$

The compensated PSM \hat{A} is then formed from the elements of $\hat{\underline{a}}$.

3.3 COMBINING ANTENNA RANGE AND RUNWAY DISTORTION MEASUREMENTS

When antenna is used as part of a larger system, which includes transmit and receive waveguide components such as circulators and mixers, the combined polarimetric distortion of the

entire system is affected by both the antenna and transmit/receive subsystems. In this section, we describe how the measurements on the antenna subsystem described in section 2, can be combined with complete system measurements (which we perform on the runway, since our radar is aircraft-mounted), can be combined to estimate the combined two-way distortion matrix. Once this matrix is found, its effects can be compensated for using the techniques described in section 3.2.

Denote the distortion matrix measured at the antenna range by

$$G(\theta, \phi) = D_{R1} D_A(\theta, \phi) \quad (3.11)$$

where D_{R1} denotes the distortion matrix corresponding to the transmitter/receiver of the compact range instrumentation, transmit waveguide runs, receiver waveguide switch, waveguide runs to mixer, and mixer; and $D_A(\theta, \phi)$ denotes the distortion matrix of the ADTS antenna itself. Note that D_{R1} is not a function of θ or ϕ ; we will also assume that D_{R1} is diagonal (i.e., neither the transmit nor the receive plumbing causes polarization leakage between the two channels).

Define for later use the diagonal matrices associated with G and D_A : Λ_G has the same diagonal elements as $G(0,0)$, but its off-diagonal elements are zero; similarly for Λ_A and $D_A(0,0)$. Note that the diagonal matrices are defined only on boresight, where $\theta = 0$, $\phi = 0$.

Since D_{R1} is diagonal, it can easily be shown that

$$\Lambda_G = D_{R1} \Lambda_A \quad (3.12)$$

Now consider the measurements made on the runway. Denote the combined distortion matrix for the entire radar system as

$$H(\theta, \phi) = D_{R2} D_A(\theta, \phi) \quad (3.13)$$

where D_{R2} denotes the distortion matrix corresponding to the ADTS transmitter/receiver used with the runway setup, and we note that the same antenna distortion matrix $D_A(\theta, \phi)$ is used since the same antenna is used. Again, D_{R2} is diagonal and independent of scanning angle. Again, denote the diagonal matrix associated with H on boresight as Λ_H .

Since D_{R2} is diagonal, it is again easily shown that

$$\Lambda_H = D_{R2} \Lambda_A \quad (3.14)$$

Combining the above results,

$$\begin{aligned}
H(\theta, \phi) &= D_{R2} D_A(\theta, \phi) \\
&= D_{R2} D_{R1}^{-1} G(\theta, \phi) \\
&= [\Lambda_H \Lambda_A^{-1}] [\Lambda_A \Lambda_G^{-1}] G(\theta, \phi) \\
&= [\Lambda_H \Lambda_G^{-1}] G(\theta, \phi)
\end{aligned} \tag{3.15}$$

Thus, the actual distortion matrix $H(\theta, \phi)$ at any scanning angles (θ, ϕ) can be inferred by combining scanning measurements Λ_G and $G(\theta, \phi)$ at the antenna range with boresight gain/phase measurements Λ_H on the runway.

4. TWO-WAY PROPERTIES OF THE ADTS ANTENNA

Section 2 described the one-way antenna polarization distortion properties, including crosspolarization. Sections 3.1 and 3.2 described how to relate one-way distortion matrices to two-way distortion matrices. In this section, we will interpret the two-way distortion matrix for the ADTS antenna in terms of worst-case crosspolarization, and show that the ADTS system specification is met without compensation. (The other forms of distortion, amplitude and gain mismatch between channels, must still be corrected for, but this is much more easily accomplished.)

The worst-case two-way crosspolarization of an antenna is approximately 6 dB worse than the worst-case one-way crosspolarization. This can be shown by the following simple argument. It is shown in Appendix A that the antenna transmit and receive distortion matrices B and C are equal. Let us ignore the gain/phase errors so that we can examine crosspolarization more closely. That is, let

$$B = C = \begin{bmatrix} 1 & \epsilon_1 \\ \epsilon_2 & 1 \end{bmatrix} \quad (4.1)$$

so that the Kronecker product yields the two-way distortion matrix D as

$$D = \begin{bmatrix} 1 & \epsilon_2 & \epsilon_2 & \epsilon_2^2 \\ \epsilon_1 & 1 & \epsilon_1 \epsilon_2 & \epsilon_2 \\ \epsilon_1 & \epsilon_1 \epsilon_2 & 1 & \epsilon_2 \\ \epsilon_1^2 & \epsilon_1 & \epsilon_1 & 1 \end{bmatrix} \quad (4.2)$$

If the target is a trihedral with $\underline{a} = [1 \ 0 \ 0 \ 1]^T$, then since $\underline{e} = D\underline{a}$, the measured polarization scattering vector \underline{e} is

$$\begin{aligned} \underline{e} &= D\underline{a} \\ &= \begin{bmatrix} 1 + \epsilon_2^2 \\ \epsilon_1 + \epsilon_2 \\ \epsilon_1 + \epsilon_2 \\ 1 + \epsilon_1^2 \end{bmatrix} \approx \begin{bmatrix} 1 \\ \epsilon_1 + \epsilon_2 \\ \epsilon_1 + \epsilon_2 \\ 1 \end{bmatrix} \end{aligned} \quad (4.3)$$

i.e., \underline{e} is merely the sum of the first and last columns of D , with VH and HV crosspolarization each equal to $\epsilon_1 + \epsilon_2$. If these two terms are of equal amplitude and phase, then their voltage sum is twice as large (i.e. 6 dB larger) than either of the summands. This turns out to be the worst case two-way crosspolarization.

Since the two-way crosspolarization is no more than 6 dB worse than the one-way crosspolarization, and since the one-way crosspolarization was reported in Section 2 to be -37 dB on

boresight and -29.5 dB within the 3 dB beam, we conclude that the two-way crosspolarization is no worse than -31 dB on boresight and -23.5 dB within the 3 dB beam. This meets specification without the need for additional polarimetric compensation to remove crosspolarization.

A. N-PORT THEORY APPLIED TO DUAL-POLARIZED ANTENNAS

In this Appendix we will model a dual-polarized antenna as a four-port system. In order to do this, some basic results from N-port theory [4] are first recalled.

A.1 N-PORT THEORY

An N-port device (see Figure A1) can be characterized by a scattering matrix which relates scattered to incident waves. At port m , the total electric field can be expressed as $E_m = (a_m + b_m)\hat{e}_m$, and the total magnetic field as $H_m = (a_m - b_m)\hat{h}_m$, where \hat{e}_m and \hat{h}_m are unit vectors for electric fields at the m -th port, a_m is the complex coefficient of an inward wave at the m -th port, and b_m is the complex coefficient of an outward wave at the m -th port. Note that both inward and outward waves are defined with respect to a common two-dimensional coordinate system (\hat{e}_m, \hat{h}_m). The scattering matrix S relates these coefficients as follows:

$$\underline{b} = S\underline{a} \quad (\text{A} - 1)$$

$$\text{where } \underline{b} = \begin{bmatrix} b_1 \\ \vdots \\ b_N \end{bmatrix} \quad \text{and} \quad \underline{a} = \begin{bmatrix} a_1 \\ \vdots \\ a_N \end{bmatrix} \quad (\text{A} - 2)$$

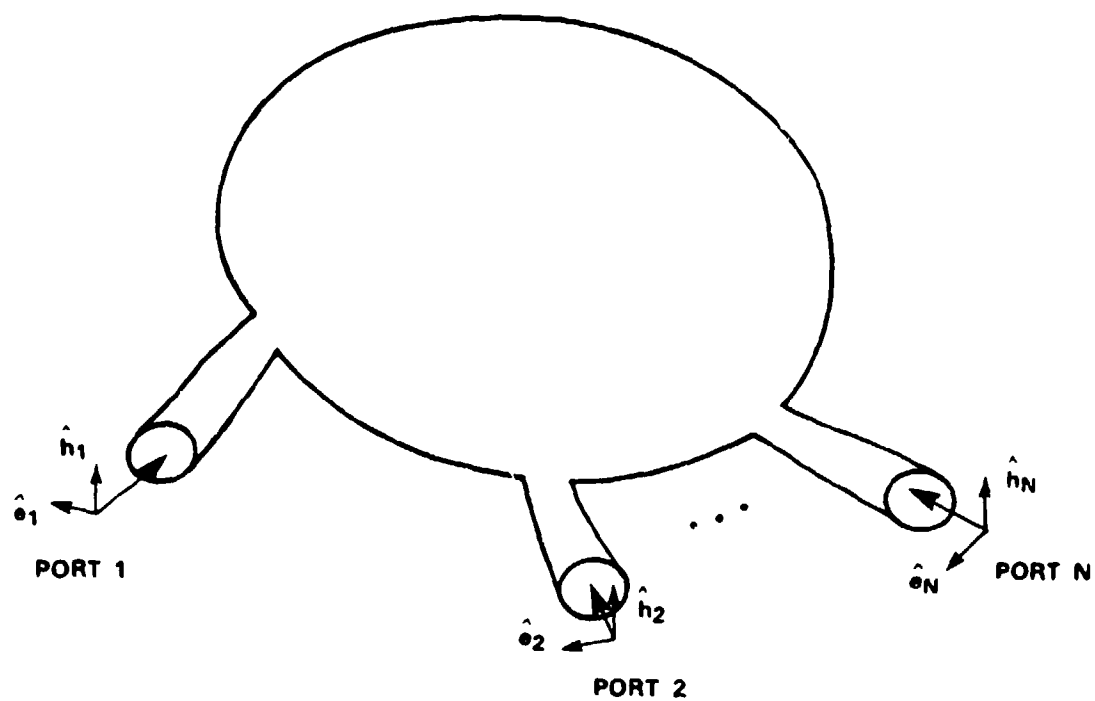
where the element S_{ij} is the amplitude of an outward wave at port i resulting from a unit inward wave at port j , with all other ports "matched" (i.e. with no other inward waves). In this expression, \underline{a} and \underline{b} are, in general, complex N-vectors.

An important result of N-port theory [4] is that when an N-port system is passive and reciprocal, the scattering matrix S is symmetric; i.e. $S_{ij} = S_{ji}$.

A.2 THE DUAL-POLARIZED ANTENNA AS A FOUR-PORT

A dual-polarized antenna has two waveguide inputs (into an orthomode transducer) and two possible axes for transmitted or received electric fields; thus it can be modelled as a four-port device.

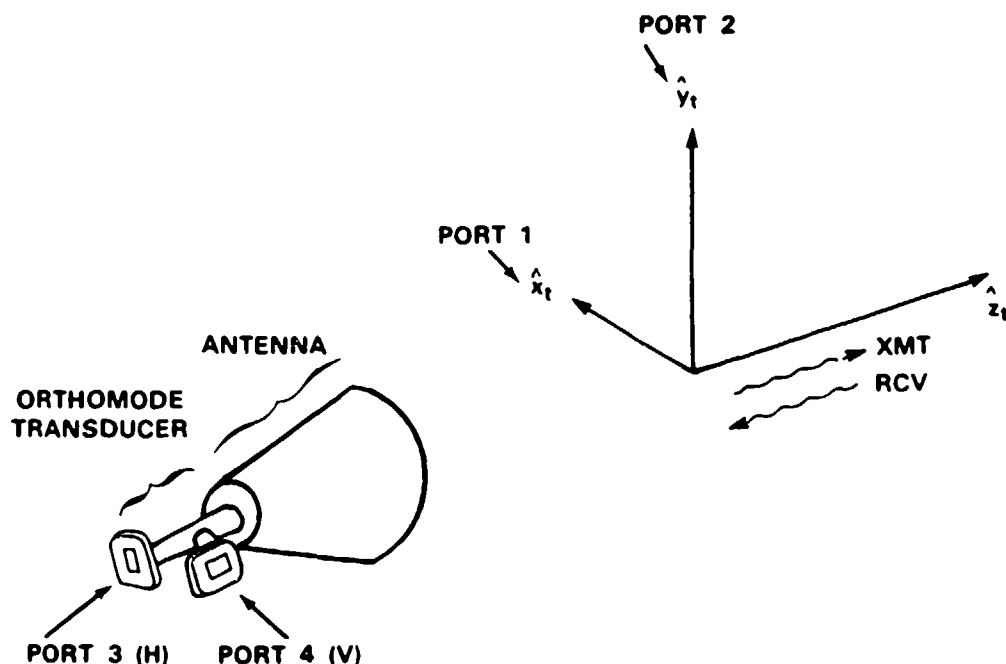
Define Port 1 and Port 2 as corresponding to horizontally- and vertically-polarized electric fields, respectively; then the unit vectors \hat{e}_1 and \hat{e}_2 (i.e., \hat{e}_m for $m = 1, 2$ referred to in section A.1) correspond to, for example, \hat{x}_t and \hat{y}_t , respectively, in Figure A2. The particular choice for \hat{e}_1 and \hat{e}_2 is not important; what is important is the fact that the same coordinates are used for the electric and magnetic field vectors for both inward and outward waves (i.e. transmitted



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Figure A1. N-port Coordinate definitions..

and received fields). Since the propagation vectors for transmitted and received fields are in opposite directions, this means that incident and scattered coordinate systems must have opposite handedness. We arbitrarily choose a right-handed triplet for transmit and left-handed for scattered, as in Figure A2.



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Figure A2. 4-port Dual-polarized Antenna Geometry..

Define Port 3 as the nominally horizontally polarized waveguide output of the orthomode transducer, and Port 4 as the nominally vertically polarized output.

Consider the operation of the antenna as a transmitter. Assume there is a unit incoming wave at Port 3 (i.e. the signal to be transmitted coming into the orthomode transducer H input). Assume also that there are no waves incident on the antenna from the rest of the world, so that the incoming waves in Ports 1, 2 and 4 are zero. The \underline{a} -vector, which is the complex four-vector describing the coefficients of the incoming waves to the four ports of the antenna, is therefore equal to $[0 \ 0 \ 1 \ 0]^T$, so that the resulting \underline{b} -vector (the complex four-vector of outgoing coefficients) is the third column of the S matrix, the elements of which are represented by S_{ij} :

$$\underline{b} = S\underline{a} = \begin{bmatrix} S_{13} \\ S_{23} \\ S_{33} \\ S_{43} \end{bmatrix} \quad (\text{A-3})$$

But the first two elements of \underline{b} are the Port 1 and Port 2 outputs; i.e. the horizontal and vertical components of the field which this antenna transmits, when the nominal transmit polarization is horizontal. Call these C_{11} and C_{21} . (This notation is consistent with that of the referenced calibration paper [1]: the electric field actually transmitted (when the nominal polarization is horizontal) is $\underline{C}_1 = \begin{bmatrix} C_{11} \\ C_{21} \end{bmatrix}$.) Thus, $S_{13} = C_{11}$ and $S_{23} = C_{21}$.

In a similar way, we can show that when the nominal transmitted polarization is vertical (i.e. into Port 4), the actual transmitted field is $\underline{C}_2 = \begin{bmatrix} C_{12} \\ C_{22} \end{bmatrix}$; since these two coefficients are equivalent to the first two elements of the fourth column of S , we have $S_{14} = C_{21}$ and $S_{24} = C_{22}$.

What has been shown so far is that the upper-right two-by-two submatrix of the four-by-four scattering matrix is the transmit distortion matrix $C = [\underline{C}_1 \underline{C}_2]$ defined in Section 3.1.

Now consider the operation of the antenna as a receiver. Assume an incident wave with horizontal component e_1 and vertical component e_2 arrives at Ports 1 and 2, and that no signals are coming into Port 3 or Port 4 (the orthomode transducer waveguide ports) from the receiver. Thus, the incoming wave vector to be used with the four-port is $\underline{a} = [e_1 \ e_2 \ 0 \ 0]^T$. Let us examine the value of the outgoing wave at Port 3; i.e. the antenna output voltage in the horizontal orthomode transducer port. This is the third component of the \underline{b} -vector, which is equal to

$$\begin{aligned} V = b_3 &= S_{31}a_1 + S_{32}a_2 + S_{33}a_3 + S_{34}a_4 \\ &= S_{13}a_1 + S_{23}a_2 + S_{33}a_3 + S_{43}a_4 \\ &= C_{11}e_1 + C_{21}e_2 \\ &= \underline{C}_1^T \underline{\hat{e}} \end{aligned} \quad (\text{A-4})$$

Similarly, the output voltage in Port 4 (vertical) can be shown to be $V = b_4 = \underline{C}_2^T \underline{\hat{e}}$.

The complex quantities b_3 and b_4 are the H and V components of the apparent measured receive field, (i.e. the distorted field). If these components are arranged to form a complex two-vector, the result is $\underline{\hat{e}}' = \begin{bmatrix} b_3 \\ b_4 \end{bmatrix} = C^T \underline{\hat{e}}$.

But the relationship $\underline{\hat{e}}' = \begin{bmatrix} b_3 \\ b_4 \end{bmatrix} = B^T \underline{\hat{e}}$ defines the receive distortion matrix B . Thus, we have shown that a dual polarized antenna (with orthomode transducer) has $B = C$; i.e. the

transmit and receive distortion matrices are equal. This result is used in Section 4 to calculate the worst-case two-way crosspolarization of an antenna.

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<p>A procedure is described for obtaining the two-way polarimetric properties of a SAR antenna from one-way measurements in a compact antenna test range. The two-way properties are determined by invoking reciprocity and computing a polarization distortion matrix (PDM). Next, one-way antenna-only measurements are combined with bore-sight measurements of the complete system to find an overall PDM. Polarimetric distortions caused by the antenna can be compensated using the inverse of the PDM; however, the Advanced Detection Technology Sensor antenna was found to have adequately low crosspolarization without the need for crosspolarization compensation.</p>					
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22a. NAME OF RESPONSIBLE INDIVIDUAL Lt. Col. Hugh L. Southall, USAF				22b. TELEPHONE (Include Area Code) (617) 981-2330	
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